



APPLICATION NO. 10/806,016

INVENTION: Multi-scale code division frequency/wavelet multiple  
access

INVENTORS: Urbain Alfred von der Embse

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will read.

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This patent application is a continuation in part of  
application 09/826,118 filed on 01/09/2001, application  
09/526,117 filed on 01/09/2001, and application 10/266,257 filed  
on 10/08/2002.

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## BACKGROUND OF THE INVENTION

### I. Field of Invention

The present invention relates to both orthogonal frequency  
15 division multiple access OFDMA, to orthogonal Wavelet division  
multiple access OWDMA, to code division multiple access CDMA, and  
to multi-scale code division multiple access MS-CDMA, for  
cellular telephone and wireless data communications with data  
rates up to multiple T1 (1.544 Mbps), E1 (2.048 Mbps), Sonet,  
20 Ethernet, and higher (>10 Gbps), and to optical CDMA and optical  
OWDMA. Applications are to wire, wireless local area, wide area,  
mobile, point-to-point, and satellite communication networks.  
More specifically the present invention relates to a new and  
novel means for combining MS-CDMA with OFDMA, to a new and novel  
25 OWDMA which is an orthogonal multi-resolution complex Wavelet  
multiple access generalization of OFDMA, and to a new and novel  
means for combining MS-CDMA with OWDMA. This new architecture  
MS-CDMA OFDMA/OWDMA is an attractive candidate to replace current  
and future OFDMA applications and CDMA applications.

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### II. Description of Related Art

Current OFDMA art is represented by the applications to the  
wireless cellular communications standards IEEE 802.11a, IEEE

802.11g, IEEE 802.15.3a, IEEE 802.16. OFDMA uses the Fourier transform basis vectors as the orthogonal channelization vectors for communications with each basis vector multiplied by a symbol which is encoded with a data or pilot signal word.

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The discrete Fourier transform DFT implemented as the fast Fourier transform FFT is defined in equations (1). Step 1 defines the digital sampling interval T over time, the sampling instants  $t=iT$  where i is the time index and where the sampling time  $1/T$  is at least equal to the complex Nyquist sampling rate to prevent spectral foldover. Step 2 is the FFT of the complex baseband transmitted signal  $z(i)$  for the data block and step 3 defines the  $N \times N$  orthogonal complex DFT matrix E row vectors  $E(k)$  which are the DFT harmonic vectors or basis vectors or code vectors or channelization vectors. Step 4 defines  $z(i)$  for one data block and is equal to the inverse FFT transform  $FFT^{-1}$  of the user symbols  $x(k)$ .

#### Unweighted DFT encoding (1)

20

##### 1 Sampling interval of DFT

	----- -----	-	-	-	-	----- -----
t →	0	T	.	.	.	(N-2)T    (N-1)T
i →	0	1				N-1    N-1

where  $1/T \geq$  complex Nyquist sample rate

25

##### 2 FFT of $z(i)$

$$\begin{aligned}
 X(k) &= FFT[ z(i) ] \\
 &= \sum_i E(k,i) z(i) \\
 &= \sum_i \exp(-j2\pi ki/N) z(i)
 \end{aligned}$$

30

##### 3 DFT orthogonal harmonic code matrix E

E =  $N \times N$  DFT orthogonal DFT matrix  
 =  $[E(k,i)]$  matrix of elements  $E(k,i)$   
 $E(k,i) = \exp(-j2\pi ki/N)$  harmonic k, time index i  
 $E(k)$  = harmonic k basis (code) vector

$$= [E(k,0), E(k,1), \dots, E(k,N-1)]$$

$$EE^* = N I$$

$$E(k)E^*(k') = \delta(k-k') N I$$

where  $I = N \times N$  identify matrix

$E^*$  = complex conjugate transpose of  $E$

$\delta(k-k') =$  Dirac delta function

$= 1$  for  $k=k'$

$= 0$  otherwise

#### 4 Transmitted DFT complex baseband signal $z(i)$ for one data block

$d(k)$  = data modulation for user  $k$

= encoded amplitude  $A(k)$  and phase  $\phi(k)$

$x(k)$  = transmitted symbol encoded with  $d(k)$

=  $A(k) \exp(j\phi(k))$

$z(i)$  =  $\text{FFT}^{-1}[x(k)]$

=  $N^{-1} \sum_k x(k) E^*(k,i)$

OFDMA for IEEE 802.11g in reference [1] is illustrated in FIG. 1. The channelization filter  $h(f)$  1 covers a 20 MHz frequency band 2 assigned to OFDMA. Plotted is the power spectral density  $\text{PSD} = |h(f)|^2$  of this channelization filter  $h(f)$ . A  $N=64$  point fast Fourier transform FFT covers this band 2. Consistent with the IEEE specification, FIG. 1 refers to the DFT which is identical to the analog Fourier transform FT since it is the sampled data version of the FT. It is convenient to consider the DFT in this invention disclosure as the digital format for the FFT. The DFT frequency spectrum 3 consists of  $N=64$  equally spaced filters 4 across this 20 MHz band. Filter spacing is equal to the DFT output rate  $1/NT = 0.3125 \text{ MHz} = 20\text{MHz}/64$ . The DFT time pulse  $p(t)$  5 is  $NT=3.2\mu\text{s}$  in length and the total DFT period 6 is  $4.0\mu\text{s}$  which allows a  $0.8\mu\text{s}$  guard time for  $p(t)$ .

Throughout this invention disclosure it will be understood that the FFT fast algorithm will always be used to implement the DFT and the inverse  $\text{FFT}^{-1}$  fast algorithm will always be used to implement the inverse  $\text{DFT}^{-1}$ .

5

OFDMA transmitter encoding of the OFDMA waveform in FIG.1 is defined in equations (2). Step 1 lists the parameters and definitions and step 2 defines the time domain weighting. Step 3 is the complex baseband transmitted signal  $z(i)$ .

10

OFDMA encoding for transmitter (2)

1 Parameters and definitions from 1 in equation (1)

and

15

$h(i)$  = 20 MHz band filter impulse response

$p(i)$  = impulse response of the DFT waveform  
= real weighting function in 6 in FIG. 1

$N$  = 64 point DFT

$1/T$  = 20 MHz sample rate for DFT  
 $\geq$  complex Nyquist rate

20

$NT$  =  $3.2\mu\text{s}$  DFT length

$1/NT$  = 0.3125 MHz DFT output rate  
= DFT channel separation  
= DFT tone spacings

52 channels are used: 4 pilot, 48 data

25

12 guard band channels for rolloff of the  $h(k)$

## 2 Pulse $p$ and band filter $h$ weighting for DFT basis vectors

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$p$  =  $p \circledast h$ , convolution of  $p$  and  $h$   
= filter transfer function in time domain  
for the combined  $h$  and  $p$  filters  
=  $[ p(0), p(1), \dots, p(N-1) ]$

3 Transmitted OFDMA encoded baseband signal  $z(i)$   
for one data block

$d(k)$  = data modulation for user  $k$   
 $=$  encoded amplitude  $A(k)$  and phase  $\phi(k)$   
 $x(k)$  = transmitted symbol encoded with  $d(k)$   
 $= A(k) \exp(j\phi(k))$   
5  $z(i) = \text{FFT}^{-1}[p(i) x(k)]$   
 $= N^{-1} \sum_k p(i) x(k) E^*(k, i)$

OFDMA for IEEE 802.11g has the strict orthogonality of the  
 DFT(FFT) replaced by cross-correlations between the 48 channel  
 10 tones and other impacts due to the band channelization and pulse  
 weighting  $p(t)$  plus the time errors  $\Delta t$  and frequency errors  $\Delta f$   
 from synchronization errors, multi-path, propagation, and  
 terminal stresses. These impacts on orthogonality are low enough  
 to allow OFDMA to support higher values for the symbol signal-to-  
 15 noise ratio  $S/N$  in the detection band that are required for  
 higher order symbol modulations. The highest order symbol  
 modulation currently is 64 state quadrature amplitude modulation  
 64-QAM corresponding to 6 bits per symbol where  $6 = \log_2(64)$  and  
 $\log_2(x)$  is the logarithm to the base 2. With rate  $3/4$   
 20 convolutional coding the highest information rate is 4.5  
 bits/symbol =  $6 \times 3/4$ . Required  $S/N$  at a  $\text{BER} = 1.0 \times 10^{-6}$  is  
 approximately  $S/N \sim 19$  dB.

OFDMA for IEEE 802.11g provides 48 channels over a 20 MHz  
 25 frequency band at a symbol rate equal to  $0.25 \text{ MHz} = 1/4.0 \mu\text{s}$  from 6  
 in FIG. 1 and for a maximum information rate equal to 4.5  
 bits/symbol this equals a burst rate of  $54 \text{ MBps} = 4.5 \times 48 \times 0.25$ .  
 Some spread spectrum properties are realized by hopping the 20  
 MHz band, shuffling the channel assignments over the 48 available  
 30 channels for a user assigned to several channels in order to  
 spread his transmissions over the band, and for "flash" OFDMA by  
 a random hopping of each user channel across the 48 available  
 channels within the band.

OFDMA receiver decoding of the OFDMA waveform in FIG.1 is defined in equations (3) for the receiver and derives estimates of the transmitted symbols by implementing matched filter detection in the receiver.

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OFDMA decoding for receiver (3)

OFDMA decoding **for one data block** derives estimates

$\hat{x}(k)$  of  $x(k)$  from the receiver estimates  $\hat{z}(i)$  of  $z(i)$

$$\hat{x}(k) = \text{FFT}[\hat{z}(i) \odot p]$$

10

$$= \sum_i \hat{z}(i) \odot p_E(k, i)$$

$$\cong x(k)$$

Current CDMA spread spectrum art is illustrated by the waveform in FIG. 2 which describes the waveform for full band CDMA communications over the band B 9 which is the output range 8 of the band filter  $h(f)$  7. The CDMA chip rate  $1/T_c$  12 is less than the available frequency band B to allow the chip frequency spectrum  $p(f)$  10,11 to roll off. As defined 9 in FIG. 2 the band B and chip rate are related by equation  $B=(1+\alpha)/T_c$  where  $\alpha$  is the bandwidth expansion factor close to  $\alpha=0.25$  for high performance communications. Frequency spectrum  $p(f)$  10 for the CDMA communications is essentially equal to the representative time pulse  $p(t)$  is a square-root raised cosine pulse which can be used for high performance communications to obtain a reasonably flat spectrum with a sharp rolloff at the edges to enable the chip rate  $1/T_c$  to be reasonably close to the available frequency band B.

Chip rate  $1/T_c$  is the CDMA total symbol rate. The users could be at different data rates but this and other architectural variations do not limit the scope of this invention. Power is uniformly spread over the CDMA pulse waveform spectrum  $p(f)$ .

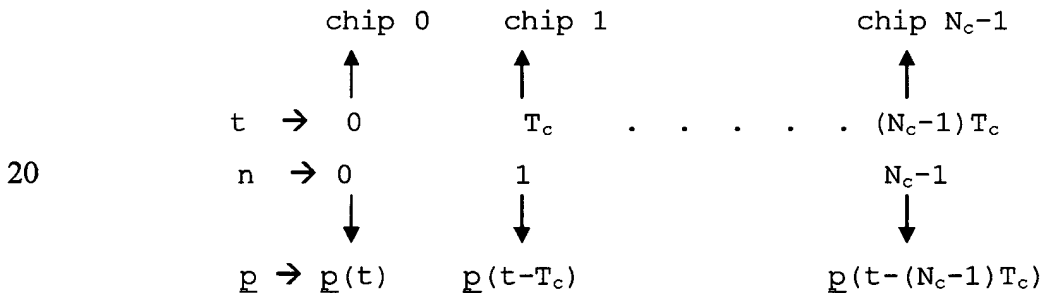
It is self evident to anyone skilled in the CDMA communications art that these communications mode assumptions are both reasonable and representative of the current CDMA art and do not limit the applicability of this invention.

5

CDMA encoding of the waveform in FIG. 2 for the transmitter is defined in equations (4). Steps 1,2 define the CDMA transmission and parameters. Step 3 defines the user symbol  $x(u)$ . Step 4 is the set of Walsh orthogonal channelization codes  $w(u)$  and step 5 is the pseudo-random PN covering or spreading code. Step 6 defines the complex baseband signal  $z(t)$  as the waveform  $p(t-nT_c)$  multiplied by the encoded sum over  $u$  and  $n$ .

15 CDMA encoding for transmitter (4)

1 CDMA transmission



25 where  $p(t-nT_c) = p(t-nT_c) \odot h(t)$   
 $=$  convolution of  $p(t-nT_c)$  and  $h$   
 $=$  filter transfer function in time domain  
for the combined  $p(t-nT_c)$  and  $h$  filters

2 Parameters and definitions

30  $N_c$  = Number of users and orthogonal code chips  
 $T_c$  = CDMA chip length or repetition interval  
 $1/N_c T_c$  = User symbol rate

3 User complex signal  $x(i)$



$d(u)$  = data modulation for user  $u$   
 $=$  encoded amplitude  $A(u)$  and phase  $\phi(u)$   
 $x(u)$  = transmitted symbol encoded with  $d(u)$   
 $= A(u) \exp(j\phi(u))$

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4 Walsh orthogonal channelization code matrix  $W$

$W$  = Code matrix,  $N_c$  rows of  $N_c$  code vectors  
 $= [W(k,n)]$  matrix of elements  $C(k,n)$

$W(u,n) = +/-1$ , chip  $n$  of vector  $u$

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$W(u)$  = code vector  $u$ , row  $k$  of  $W$   
 $= [W(u,0), W(u,1), \dots, W(u, N_c-1)]$

5 PN covering (spreading) code  $P(n)$  for chip  $n$

$P(n) = \exp(j\phi(n))$

15

6 Transmitted CDMA complex baseband signal  $z(t)$

$z(t) = N_c^{-1} \sum_u \sum_n P(n)W(u,n)x(u) p(t-nT_c)$

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CDMA decoding of the waveform in FIG. 2 for the transmitter  
 is defined in equations (5). Step 1 defines the convolution  
 $R(n, n-n')$  of the CDMA pulse waveform with itself in the matched  
 filter receiver. Steps 2,3 are the Walsh and PN decoding  
 properties. Step 4 uses the matched filter detection theorem to  
 derive the estimates of the transmitted symbols.

25

CDMA decoding for receiver (5)

1 Parameters and definitions are defined in

1,2,3 in equations (4) together with

30

$R(n, n-n')$  = convolution of  $p(t-nT_c)$  with  
 $p(t-n'T_c)$  evaluated at the receiver  
 detection times  $t=nT_c$   
 $= R(n) \delta(n-n')$   
 $= R(n)$  for  $n=n'$

$$= 0 \quad \text{otherwise}$$

2 Walsh decoding of channelization codes  $W(k)$

$$WW^* = N_c I$$

5 where  $I = N_c \times N_c$  identify matrix

$W^* = \text{conjugate transpose of } W$

$$\langle W(k), W(k') \rangle = N \delta(k-k')$$

3 PN decoding

10  $P(n)P^*(n) = 1$  for all  $n$

where  $P^* = \text{complex conjugate of } P$

4 CDMA decoding

$$\hat{x}(k) = \sum_n P^*(n)W^*(k,n) \hat{z}(t) \otimes p(t-nT_c)$$

15

It should be obvious to anyone skilled in the communications art that these example implementation algorithms in equations (1), (2), (3), (4), (5) clearly define the fundamental OFDMA and CDMA signal processing relevant to this invention disclosure and it is obvious that this example is representative of the other possible signal processing approaches.

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For cellular applications the encoding algorithms for the transmitter describe the implementation of OFDMA and of CDMA encoding and are the transmission signal processing applicable to this invention for both the hub and user terminals, and the decoding algorithms for the receiver describes the corresponding OFDMA and CDMA receiving signal processing for the hub and user terminals for applicability to this invention.

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For optical communications applications the microwave processing at the front end of both the transmitter and the receiver is replaced by the optical processing which performs the complex modulation for the optical laser transmission in the

transmitter and which performs the optical laser receiving function of the microwave processing to recover the complex baseband received signal with the remainder of the signal processing functionally the same for the OFDMA and for the CDMA encoding transmitter and functionally the same as described for the OFDMA and CDMA receiving signal processing receiver.

## BRIEF SUMMARY OF THE INVENTION

This invention introduces the new OWDMA communications technology which implements orthogonal multi-resolution complex Wavelet division multiple access and is a multi-resolution complex Wavelet generalization of OFDMA; introduces the new application of the multi-scale code division multiple access MS-CDMA architecture which integrates MS-CDMA with OFDMA and with OWDMA, and introduces the variable gain control over frequency.

The new OWDMA forms a uniform set of contiguous orthogonal filters across the available frequency band in one of the numerous available architectural options and which implements a polyphase filter bank across the available frequency band with the basic property that the filters are orthogonal. Each filter defines a OWDMA channel for communications. Similar to OFDMA, the symbol rate within each channel is equal to the channel separation  $1/NT$  where  $N$  is the number of channels,  $1/T$  is the frequency band, and  $NT$  is the symbol-to-symbol separation and which is made possible by using multi-resolution complex Wavelet channelization filters developed in reference [2]. OWDMA filters are orthogonal in frequency which means their frequency spectrums are non-overlapping, and they have flat spectrums across each channel.

OWDMA orthogonality only requires frequency synchronization whereas OFDMA orthogonality requires both time and frequency synchronization which means that time synchronization errors on the user communication channels for the return links to the hub or access point for cellular communications do not degrade the orthogonal separation between the channels as they will for OFDMA.

Sensitivity to frequency synchronization errors is less for OWDMA primarily because of the non-overlapping of the frequency spectrums. The overall tolerance to user-to-user imbalances on the return communications channels is better for OWDMA primarily because their design keeps the frequency spectrums from overlapping in the presence of real operational conditions with synchronization errors. This allows the forward communications link to support a power imbalance between channels to mitigate differing ranges and path losses at the user antennas.

Application of multi-scale code division multiple access MS-CDMA in reference [2] provides a means to implement the new hybrid-Walsh orthogonal CDMA codes in reference [4] over the OFDMA/OWDMA channels by spreading the CDMA within each channel and over all of the channels such that each user can be spread over the complete band. This keeps the chip rate equal to  $1/NT$  while maintaining the spreading over the fullband  $1/T$  and allows the band transmit Tx power to be independently controlled in frequency. These two CDMA scales  $1/NT, 1/T$  are generated by MS-CDMA in combination with the OFDMA/OWDMA channelization filter banks. The  $1/T$  scale is to combat fading and interference similar to the current CDMA, and the  $1/NT$  scale is for acquisition, synchronization, and equalization protection against multi-path and provides the flexibility for band power control to provide a frequency diversity communications improvement.

Variable control over the frequency B can be implemented by partitioning the CDMA over the channels into separate groups and assigning an independent power level to each group of channels. OWDMA readily supports differences in power levels between adjacent channels. Power control is desirable to support differences in quality of service SoC, range losses, and path loss.

It should be obvious to one familiar with the CDMA communications art that the number of scales could be a larger number than the two used in this invention disclosure and the multi-resolution complex Wavelet design for OWDMA supports the partitioning of the frequency band  $1/T$  into several frequency scales for the channelization filters, supports individual multi-resolution complex Wavelet packets for communications in the time-frequency domain and which can be integrated into the MS-CDMA, and supports separate and segmented communications bands simultaneously.

## BRIEF DESCRIPTION OF SEVERAL VIEWS OF THE DRAWINGS

The above-mentioned and other features, objects, design algorithms, and performance advantages of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings and performance data wherein like reference characters and numerals denote like elements, and in which:

FIG. 1 is a description of the OFDMA waveform in the frequency and time domains over a 20 MHz band for the IEEE 802.11g standard.

FIG. 2 is a description of the CDMA waveform in the frequency and time domains over a band B.

FIG. 3 is a description of the OWDMA waveform in the frequency and time domains over a band B.

FIG. 4 is a description of the design requirements on the multi-resolution complex Wavelet  $PSD=|\psi(f)|^2$ .

FIG. 5 is a representative baseband frequency response for the multi-resolution complex Wavelet filter and the square-root cosine filters for the bandwidth expansion factors =0.22, 0.40.

FIG. 6 is a representative implementation of MS-CDMA OFDMA/OWDMA encoding for transmitter.

FIG. 7A is a representative MS-CDMA OFDMA/OWDMA transmitter implementation block diagram. the.

FIG. 7B is a representative MS-CDMA OFDMA/OWDMA transmitter MS-CDMA mapping.

FIG. 8 is a representative implementation of MS-CDMA OFDMA/OWDMA encoding for the receiver.

FIG. 9 is a representative implementation block diagram for the MS-CDMA OFDMA/OWDMA receiver.

## DETAILED DESCRIPTION OF THE INVENTION

This invention introduces the new orthogonal multi-resolution complex Wavelet division multiple access OWDMA communications technology which is a multi-resolution complex Wavelet generalization of OFDMA, introduces the new multi-scale

code division multiple access MS-CDMA architecture to integrate MS-CDMA with OFDMA, and introduces the new MS-CDMA architecture to integrate MS-CDMA with OFDMA,

5        **OWDMA uses the multi-resolution complex Wavelet waveform**  
developed in reference [2] to generate multi-rate orthogonal  
filter banks and waveforms to support communications at the  
critical symbol rate equal to a combined  $1/T$  symbols per second  
for a  $1/T$  Hz frequency band. The critical symbol rate is the  
10 Nyquist sample rate and corresponds to no excess bandwidth  $\alpha=0$  in  
the equation  $B=(1+\alpha)/T$  which can be rewritten as the bandwidth-  
time product  $BT=(1+\alpha)$  that relates filter bandwidth  $B$  and the  
symbol rate  $1/T$  for communications supported by this filter. For  
the OFDMA filter in FIG. 1 it is observed  $BT=1+\alpha=64/52=1.33$   
15 corresponding to  $\alpha=0.33$  since 52 of the 64 DFT filters generated  
are used and for the CDMA filter in FIG. 2 it is observed that  
the bandwidth-time product is about  $BT=1.25$  corresponding to  
 $\alpha=0.25$  for the 3G CDMA cellular communications systems.

20        Multi-resolution complex Wavelet design algorithms were  
developed in reference [2] as a means to design polyphase  
multirate filters, quadrature mirror filters (QMF), perfect  
reconstruction filters, Wavelet iterated filter banks, and  
Wavelet tiling of the time-frequency t-f space. Prior to the  
25 invention of multi-resolution complex Wavelet design algorithms,  
theoretical studies had not yielded useful realizable filters  
for system applications implementing these architectures as  
summarized by the digital filtering and polyphase research in  
references [5],[6] and the Wavelet research as summarized in the  
30 references [7],[8],[9],[10],[11].

Multi-resolution complex Wavelets defined in equations (6),  
(7),(8) expand the Wavelet analytical formulation to include a  
frequency variable which specifies the center frequency of our

new Wavelets and performs a frequency translation of the real mother Wavelet to this center frequency. Currently, Wavelets are real functions of the scale and translation parameters. Multi-resolution complex Wavelets are functions of these parameters plus the frequency variable. This new concept of frequency as an additional parameter provides an added degree of flexibility and together with the Fourier domain design approach provide an entirely new means for deriving these new waveforms as a generalization of the traditional real Wavelets and their modification to include the added frequency parameter. With frequency translation as the additional parameters the analytical formulation of these new waveforms as a function of the baseband or mother waveform centered at dc (dc refers to the origin  $f=0$  of the frequency space) is given in equations (6).

Equations (6) introduce the new multi-resolution complex Wavelets. Step 1 in equations (6) is the definition for the continuous real Wavelet over the time-frequency  $t$ - $f$  space from reference [7],[9],[12] where  $\psi(t)$  is the Wavelet which is a waveform of finite extent in time  $t$  and frequency  $f$  over the  $t$ - $f$  space. Wavelet parameters  $a, b$  are the Wavelet dilation and translation respectively or equivalently are the scale and shift. The  $\psi(t)$  without the indices  $a, b$  is the mother Wavelet which is a real and symmetric localized function in the  $t$ - $f$  space used to generate the doubly indexed Wavelet  $\psi(t|a, b)$  where  $\psi(t|a, b)$  reads the  $\psi$  is a function of time  $t$  for the parameters  $a, b$ . Scale factor  $|a|^{-1/2}$  has been chosen to keep the norm of the Wavelet invariant under the parameter change  $a, b$ . Norm is the square root of the energy of the Wavelet response. The Wavelets  $\psi(t|a, b)$  and  $\psi(t)$  are localized functions in the  $t$ - $f$  space which means that both their time and frequency lengths are bounded.



Step 2 in equations (6) defines the digital Wavelet which is a Wavelet over the digital t-f space corresponding to digitization of the t-f space at the  $1/T$  analog-to-digital A/D rate where  $T$  is the interval between digital samples and "i" is the index over the digital samples. Time "t" for the continuous Wavelet in 1 is replaced by the equivalent digital sample number "i" corresponding to  $t=iT$  at sample i. Wavelets in digital t-f space have an orthogonal basis that is obtained by restricting the choice of the parameters a,b to the values  $a=2^{-p}$ ,  $b=qN2^p$  where p,q are the new scale and translation parameters and N is

New multi-resolution complex Wavelets (6)

1 Current continuous Wavelet as a function of the mother Wavelet at dc

$$\psi(t|a,b) = |a|^{-1/2} \psi((t-b)/a)$$

where a,b are the dilation, shift parameters

2 Current digital real Wavelet

Digital Wavelet shift q and scale p

$$a = 2^p$$

$$b = q N 2^p$$

Digital Wavelet as a function of mother Wavelet

$$\psi(i|p,q) = 2^{-p/2} \psi(2^{-p} i - qN)$$

where N = number of samples over the Wavelet

spacing or repetition interval

$T_s = NT$  wavelet spacing

i = digital sampling index

T = digital sampling interval

t = iT time index at digital sample i

3 Our new digital multi-resolution complex Wavelet is a function of the mother Wavelet at dc

3.a) Generalized complex format

$$\psi(i|p,q,k) = 2^{-p/2} \psi(2^{-p} i - qN) \exp[j2\pi f_c(p,k)2^{-p}iT]$$

3.b) Multi-rate filter complex format

$$\psi(i|p,q,k) = 2^{-p/2} \psi(2^{-p} i - qN) E^*(k2^p, 2^{-p} i)$$

4 Orthogonality equations for our new digital  
5 complex Wavelet

$$\sum_i \psi(i|p,q,k) \psi^*(i|p',q',k') = N \quad \text{iff } p=p', q=q', k=k' \\ = 0 \quad \text{otherwise}$$

5 New digital complex Wavelet for a uniform filter bank  
as a function of the mother Wavelet at dc

10  $\psi(i|p=0,q,k) = \psi(i-qN) E^*(k,i)$

the spacing or repetition interval  $T_s=NT$  of the Wavelets (which  
from a communications viewpoint are symbols) at the same scale  $p$ .  
Wavelets at  $p,q$  are related to the mother Wavelet by the equation  
15 in 2 where the mother Wavelet is a real and even function of the  
sample coordinates and which follows directly from 1 for the  
continuous  $t$ - $f$  space.

Steps 3,4 in equations (6) define our new Wavelet in the  
20 digital  $t$ - $f$  space and their orthogonality properties. Our new  
Wavelets are complex generalizations of Wavelets in  $t$ - $f$  space  
which enable them to be useful for communications and radar  
applications. This generalization is accomplished 1) by the  
addition of a frequency translation parameter  $k$  which controls  
25 the frequency offset of the Wavelet, 2) by generalizing the  
Wavelet weighted orthonormality properties in step 4 to apply to  
waveforms over the time translation  $q$  and over the scales  $p$  with  
the inclusion of the frequency translation and where  $\psi^*$  is the  
complex conjugate of  $\psi$  and is required in the orthogonality  
30 equations since the multi-resolution complex Wavelet  $\psi$  becomes  
complex with the addition of harmonic  $k$ , and 3) by their  
characterization and design in the Fourier domain. The frequency  
parameter  $k$  controls the frequency translation  $\exp[j2\pi f_c(p,k)2^{-p}iT]$  in 3 for the generalized format and the frequency

translation  $E^*(k2^p, 2^{-p}i)$  in 3 for the multi-rate filter format. With this frequency translation the analytical formulation 3 of these new Wavelets is given as a function of the baseband or mother waveform centered at dc corresponding to  $k=0$ . Purpose of the frequency index  $k$  is to identify the center frequencies of the waveforms at the scale  $p$  and time translation  $q$  in the  $t$ - $f$  space. The generalized center frequency  $f_c(p,k)$  of the frequency translated dc waveform at scale  $p$  and frequency index reduces to  $f_c(p,k) = k2^p/NT$  for application to multi-rate filters.

10

Step 5 in equations (6) defines the equations for our new Wavelet for a uniform polyphase filter bank which is one of several OWDMA candidate architectures. Our new Wavelets in 5 are the impulse responses of the corresponding digital symbols encoded with digital data for transmission which is the synthesis filter bank in polyphase theory, and are the filter bank detection impulse responses for recovery of the transmitted digital symbols by the analysis filter bank in polyphase theory assuming matched filter detection. The digital filters are observed to be the DFT's of the mother wavelets similar to the construction of the OWDMA in 3 in equations (2) upon replacing the OWDMA pulse waveform  $p$  with the mother Wavelet  $\psi(i-qN)$  in 5 and setting  $q=0$  corresponding to the symbol at  $t=0$ .

15

20

Equations (7) derive the multi-resolution complex Wavelet as a function of the Fourier design coordinates. Design algorithms provide a means to design the mother Wavelet in the Fourier frequency domain to fit the communications and radar specifications. From this mother Wavelet, the Wavelets at the appropriate scales  $\{p,q,k\}$  are easily found as demonstrated in 3,5 in equations (6). Design in the frequency domain means the design coordinates specifying the Wavelet are Fourier frequency harmonics or coordinates. Step 1 lists the parameters and

25

30

coordinates and step 2 defines the Fourier harmonic frequency design coordinate.

Step 3 is the DFT representation from equations 3 in equations (1) of the real mother Wavelet  $\psi(i)$  as a function of the DFT Fourier harmonic coefficients  $\{\psi(k')\}$ . Step 4 derives the equation for the multi-resolution complex Wavelet as a function of the DFT Fourier harmonic coefficients by substituting 3 in equations (7) into 3 in equations (6).

Digital multi-resolution complex Wavelet is a function of the Fourier harmonic design coordinates

1 Parameters and coordinates

$\psi = \psi(i)$  mother Wavelet

$= \psi(i|p=0,q=0,k=0)$

$N' =$  length of the mother Wavelet  $\psi(i)$

$= NL+1$

$N =$  sampling interval of  $\psi$

$=$  spacing of  $\psi$  for orthogonality

$L =$  length of  $\psi$  in units of  $N$

$T =$  spacing of digital samples

$T_s = N'T$  length of  $\psi$  in seconds

2 Fourier harmonic frequency design coordinates

$\psi(k') =$  value of the Fourier design harmonic  $k'/N'T$

$=$  Fourier harmonic design coordinate value

$\{k'\} =$  set of harmonic design coordinates

for  $k' = 0, +/-1, . . .$

corresponding to harmonics  $\{k'/N'T\}$

3 Multi-resolution real mother Wavelet definition

$\psi(i) = \sum_{k'} \psi(k') E^*(k', i)$

#### 4 Multi-resolution complex Wavelet definition

$$\begin{aligned}\psi(i|p,q,k) \\ = 2^{-p/2} \sum_k \psi(k') E^*(k'/L + k2^p, 2^{-p}(i - qN))\end{aligned}$$

5 Several fundamental properties follow directly from the frequency design approach in 4 in equations (7). It is demonstrated in reference [2] that our multi-resolution complex Wavelets are implemented with our design in the Fourier domain and our multi-resolution complex Wavelet design remains invariant  
10 under scale changes. It is demonstrated in reference [2] that the Fourier frequency domain design in 3 in equations (7) remains invariant for all parameter changes and in particular for all scale changes.

15 Multi-resolution complex Wavelet design algorithms are illustrated by a representative least-squares LS design for a OWDMA polyphase uniform filter bank in FIG. 3. There are two categories of LS algorithms and these are the eigenvalue and the gradient search that respectively can be reduced to algorithms  
20 equivalent to the original eigenvalue [13] and Remez-exchange [14] waveform design algorithms for application to a uniform filter bank.

FIG.3 is an example of an orthogonal multi-resolution  
25 complex Wavelet division multiple access OWDMA Wavelet polyphase filter bank over the band B by setting N in 5 in equations (6) equal to the N in 17 in FIG. 3. The band channelization filter is a roofing filter  $h(f)$  14 that covers the B frequency band 15 assigned to OWDMA. Unlike its use in FIG. 1 for the OWDMA with  
30 B=20 MHz this filter is intended to be flat over the frequency band B of interest and with a bandwidth 18 equal to  $1/T > B$ . Plotted is the power spectral density  $PSD = |h(f)|^2$  of this channelization filter  $h(f)$ . The  $h(f)$  output is digitized at the sample rate  $1/T$  to form the OWDMA multi-resolution complex

Wavelet polyphase filter bank **16** which are uniformly spaced **17** at  $1/NT$  Hz over the  $1/T$  frequency band  $1/T$ . This digital sample rate  $1/T$  is sufficiently large to allow the use OWDMA filters in **17** in FIG. 3 for communications over the band B **18** with no excess  
5 bandwidth  $\alpha=0$  unlike the OFDMA in FIG. 1 which has  $\alpha=0.33$  and CDMA in FIG. 2 with  $\alpha=0.25$ .

This OWDMA polyphase filter bank is ideally decimated which means the filter output multi-resolution complex Wavelet sample  
10 rate  $1/T_s$  is equal to the channel-to-channel spacing  $1/T_s=1/NT$  equivalent to stating that there is no excess bandwidth  $\alpha=0$  within the filter bank. A representative **19** OWDMA multi-resolution complex Wavelet  $\psi$  for  $L=8$  is plotted in **20** as a function of the time offset expressed in units of the multi-  
15 resolution complex Wavelet spacing  $NT$ . This  $\psi$  was designed by the eigenvalue category of LS design algorithms. Our design for this topology is immediately applicable to an arbitrary set of multi-resolution OWDMA filters through the scaling **3,4** in equations (7) which gives the design of our Wavelet at arbitrary scales in  
20 terms of our design of the mother Wavelet.

Multi-scale mother Wavelet  $\psi$  design for the OWDMA polyphase filter bank in FIG. 3 starts by using the frequency design template in FIG. 4 to construct the LS error metrics as a  
25 function of the DFT frequency design coordinates  $\{\psi(k')\}$  in **2** in equations (7) which define the mother Wavelet in **3** in equations (7). Next minimization search algorithms are developed and used to find the values of  $\{\psi(k')\}$  that minimize the weighted sum of these LS design error metrics equal to the cost function  $J$  in  
30 equations (8) for the LS design. Minimizing  $J$  with respect to the  $\{\psi(k')\}$  gives the design values of  $\{\psi(k')\}$  for constructing the mother Wavelet in **3** in equations (7).

Equations (8) define the LS cost function  $J$  constructed with the LS metrics with the aid of the frequency design template in FIG. 4. Step 1 defines the LS error metrics which are the passband error metric  $\mu(1)$ , stopband error metric  $\mu(2)$ , quadrature mirror filter QMF error metric  $\mu(3)$ , intersymbol interference error metric  $\mu(4)$ , and the adjacent channel interference error metric  $\mu(5)$ .

# **LS cost function $J$ for designing $\psi(i)$ (8)**

## **1 LS metrics**

$\mu(1)$  = passband LS error metric measures the LS error of the passband ripple 24 in FIG.4

$\mu(2)$  = stopband LS error metric measures the stopband attenuation 28 in FIG. 4

$\mu(3)$  = quadrature mirror filter QMF LS error metric measures the flatness of the sum of two contiguous filters over the deadband 31 in FIG. 4

$\mu(4)$  = orthogonality LS error metric measures the intersymbol interference ISI between overlapping Wavelets of different wymbol

$\mu(5)$  = orthogonality LS error metric measures the adjacent channel interference ACI from nearest neighbor channels

## **2 Metric weighting**

$w(n)$  = metric weight for error metrics  $n=1,2,3,4,5$  in 1  
 $\geq 0$

such that  $\sum_n w(n) = 1$

## **3 Cost function $J$**

$J = \sum_n w(n) \mu(n)$

FIG. 4 is the frequency design template for the power spectral density  $PSD=|\psi(f)|^2$  of the multi-resolution complex Wavelet and defines the parameters of interest for the passband metric  $\mu(1)$  and stopband metric  $\mu(1)$  in step 1 in equations (8).

5 Passband 21 of the wavelet  $PSD=|\psi(f)|^2$  is centered at dc ( $f=0$ ) since we are designing the mother Wavelet, and extends over the frequency range  $\omega_p$  extending from  $-\omega_p/2$  to  $+\omega_p/2$  22 in units of the radian frequency variable  $\omega=2\pi fT$  23 where  $T$  is the digital sampling interval defined in FIG. 3. The frequency space extends

10 over the range of  $\Delta f=-1/2T$  to  $\Delta f=+1/2T$  which is the frequency range in FIG 3 and the mother Wavelet is at the center of the frequency band. Quality of the  $PSD=|\psi(f)|^2$  over the passband is expressed by the passband ripple 24. Stopband 25 starts at the edge 26 of the passbands of the adjacent channels  $\pm\omega_a/2$  26 and

15 extends to the edge of the frequency band  $\omega=\pm\pi$  27 respectively. Stopband attenuation 28 at  $\pm\omega_a/2$  measures the  $PSD=|\psi(f)|^2$  isolation between the edge of the passband for the mother Wavelet and the start of the passband for the adjacent Wavelet channel centered at  $\pm\omega_s$  29. Rolloff 30 of the stopband

20 is required to mitigate the spillover of the Wavelet channels other than the adjacent Wavelet channels, onto the mother Wavelet channel. Deadband or transition band 31 is the interval between the passbands of contiguous Wavelet channels, and is illustrated in FIG. 4 by the interval from  $\omega_p/2$  to  $\omega_a/2$  between the mother

25 Wavelet channel and the adjacent Wavelet channel at  $\omega_a$ . Waveform sample rate  $\omega_s$  32 is the waveform repetition rate. For the LS example algorithm, the waveform sample rate is equal to the channel-to-channel spacing for zero excess bandwidth. Therefore,

30  $1/T_s = \omega_s/2\pi T = 1/NT$  which can be solved to give  $\omega_s = 2\pi/M$  for the radian frequency sampling rate of the filter bank which is identical to the Wavelet repetition rate.



Equations (8) step 1 QMF LS error metric  $\mu(3)$  expresses the requirement on the deadband that the PSD's from the contiguous channels in FIG. 4 add to unity across the deadband  $31$   $[\omega_p, \omega_s]$  in FIG. 4 in order that the Wavelets be QMF filters.

5

Equations (8) step 1 Inter-symbol interference ISI and ACI error metrics  $\mu(4), \mu(5)$  are orthonormality metrics that measure how close we are able to design the set of Wavelets to be orthonormal over the t-f space, with the closeness given by the

10 ISI error metric  $\mu(4)$  and the ACI error metric  $\mu(5)$ . ISI is the non-orthogonality error between Wavelets within the same channel separated by multiples of the sampling interval  $1/MT$  seconds where  $T$  is the sample time and  $M$  is the interval of contiguous samples. Adjacent channel interference ACI is the non-

15 orthogonality error between between Wavelets within a channel and the Wavelets in adjacent Wavelet channels at the same sample time and at sample times separated by multiples of the sample interval. As observed as noise contributions within each sample in a given channel, the ISI is the noise contribution due to the

20 other received Wavelets at the different timing offsets corresponding to multiples of the sampling interval. Likewise, the ACI is the noise contribution due to the other Wavelets in adjacent Wavelet channels at the same sampling time and at multiples of the sampling interval.

25

Equations (8) step 1 ISI and ACI errors are fundamentally caused by different mechanisms and therefore have separate metrics and weights to specify their relative importance to the overall sum of the LS metrics. ISI is a measure of the non-

30 orthogonality between the stream of Wavelets within a channel as per the construction in FIG. 4. On the other hand, ACI is a measure of the non-orthogonality between the Wavelets within a channel and the other Wavelets in adjacent channels. This means the stopband performance metric has a significant impact on the

ACI due to the sharp rolloff in frequency of the adjacent channel, and the ACI metric is then a measure of the residual non-orthogonality due to the inability of the stopband rolloff in frequency from completely eliminating the ACI errors.

5

Equations (8) step 2 defines the weights of the LS error metrics when summed to yield the cost function  $J$ . These weights are real and normalized to sum to unity. They have proven to be helpful in the Wavelet design to emphasize the relative importance of the individual error metric contributions to  $J$ .

10

Equations (8) step 3 defines the cost function  $J$  as the weighted sum of the LS error metrics and which is minimized with respect to the DFT frequency design harmonics  $\{\psi(k')\}$  to select the best LS choice for the  $\{\psi(k')\}$  to design the mother Wavelet in 3 in equations (7) and the channel Wavelets by frequency translation in 5 in equations (6).

15

Multi-scale mother Wavelet frequency response in FIG. 5 is evaluated by implementing the LS design algorithms in reference [2] that minimize the  $J$  in 3 in equations (8) to find the best set of DFT frequency design coordinates  $\{\psi(k')\}$  which give the mother Wavelet 3 in equations (7). FIG. 5 plots the PSD frequency response for the multi-scale mother Wavelet and the square-root (sq-rt) raised-cosine (r-c) waveforms with excess bandwidth  $\alpha = 0.22, 0.40$  which waveforms are extensively used for other communications. Plotted are the measured PSD in dB 42 versus the frequency offset 43 from dc expressed in units of the symbol rate. for the new Wavelet waveform 44, the sq-rt r-c with  $\alpha = 0.22$  45, and the sq-rt r-c with  $\alpha = 0.40$  46. It is believed that the multi-resolution complex Wavelets can be designed as a filter with better performance parameters than possible with any other known algorithm.

25

30

OWDMA encoding for the transmitter is defined in equations (9). Step 1 lists parameters and definitions and step 2 defines the transmitted OWDMA encoded baseband signal  $z(i)$  for contiguous data blocks and where the symbol offsets  $\Delta$  account for symbol overlaps over the symbol  $q$  data block interval within each channel.

OWDMA encoding for transmitter (9)

1 Parameters and definitions

10  $h(i)$  = roofing filter impulse response in time  
for  $h(f)$  in 15 in FIG.3

$\psi(i)$  =  $\psi(i|p=0, q=0, k=0)$  mother Wavelet in 3  
in equations (7)  
= baseband or dc  $k=0$  Wavelet at  $p=0, q=0$

15  $N$  = number of OWDMA filters over the  $1/T$   
frequency band

$1/T$  = digital sample rate for OWDMA  
 $\geq$  complex Nyquist rate for roofing filter  $h(f)$

$NT$  = OWDMA Wavelet spacing

20  $1/NT$  = OWDMA Wavelet output rate  
= OWDMA Wavelet channel separation

$NBT$  = channels used for data and pilot  
where  $B$  is the frequency band in FIG. 3

$N(1-BT)$  = guard band channels for rolloff  
of the  $h(f)$

25  $d(k, q)$  = data modulation for user  $k$  for data block  $q$   
= encoded amplitude  $A(k|q)$  and phase  $\phi(k|q)$

$x(k|q)$  = transmitted symbol encoded with  $d(k)$   
=  $A(k|q) \exp(j\phi(k|q))$

30 Assume the  $h(f)$  is flat over the passband for  
both Tx and Rx and can be neglected

2 Transmitted OWDMA encoded baseband signal  $z(i)$

Time index field referenced to  $q=0$

$i = 0, +/-1, +/-2, . . .$   
 $=$  digital sampling time index  
 $i_0 =$  index over a  $\psi(i)$  spacing interval  
 $= 0, 1, 2, . . . , N-1$   
 $\Delta =$  index over the  $\psi(i)$  range in units  
of the  $\psi(i)$  spacing interval  $N$   
 $= (L/2-1), \dots, -1, 0, +1, \dots, +(L/2-1)$

Complex baseband signal

$z(i_0|q) =$  complex baseband signal over  $i_0$   
for data block  $q$   
 $= \text{FWT}[ x(k|q+\Delta) ]$   
 $= N^{-1} \sum_{\Delta} \sum_k x(k|q+\Delta) \psi(i|q+\Delta, k)$   
 $= N^{-1} \sum_{\Delta} \sum_k x(k|q+\Delta) \psi(i_0+\Delta N) E^*(k, i_0)$   
 $z(i) = \sum_q z(i_0|q)$

where FWT is the fast multi-resolution complex  
Wavelet transform

3 FWT algorithm for OWDMA encoding in the transmitter

3.a) FWT pre-calculation  $\text{FFT}^{-1}$

$\lambda(i_0, q+\Delta) = N^{-1} \sum_k x(k|q+\Delta) E^*(k, i_0)$   
 $= \text{FFT}^{-1}[ N^{-1} \sum_k x(k|q+\Delta) E^*(k, i_0) ]$

3.b) FWT post-sum

$z(i_0|q) = \sum_{\Delta} \psi(i_0+\Delta N) \lambda(i_0|q+\Delta)$

4 Computational complexity of fast algorithm

Real multiply rate  $R_M$

$$R_M T = 2 \log_2(N) + 2 L$$

Real add rate  $R_A$

$$R_A T = 3 \log_2(N) + 2 L$$

Step 3 is the new fast FWT algorithm in this invention  
disclosure for the transmitted OWDMA which consists of the pre-

calculation  $\text{FFT}^{-1}$  in sub-step **3.a** followed by a post-sum in sub-step **3.b**.

Step **4** evaluates the real multiply complexity metric  $R_{MT}$  and  
 5 real add computational complexity metrics  $R_{AT}$  in terms of  
 multiplies/adds per digital sample for the fast algorithm in step  
**3**. The first term in these metrics is the complexity of the  $\text{FFT}^{-1}$   
 for a base 2 and the second term is the complexity of extending  
 the multi-resolution complex Wavelet waveform over  $L$  of the  
 10 symbol intervals.

OWDMA decoding for the receiver is defined in equations  
 (10). Step **1** refers to **1,2** in equations (9) for the parameters  
 and definitions and defines the OWDMA filtering Wavelet. Step **2**  
 15 demonstrates Wavelet orthogonality. Estimates of the transmitted  
 symbols in step **3** are equal to the  $\text{FWT}^{-1}$  of the received baseband  
 signal.

OWDMA decoding for receiver (10)

20 **1** Parameters and definitions are defined in **1,2** in  
 equations (9) together with

OWDMA filtering wavelet from **5** in equations (6)

$$\begin{aligned}\psi^*(i|q, k) &= \psi^*(i|p=0, q, k) \\ &= \psi(i-qN)E(k, i_0)\end{aligned}$$

25 Assume the  $h(f)$  is flat over the passband for  
 both Tx and Rx and can be neglected

**2 Multi-resolution complex Wavelet orthogonality** from **4** in  
 equations (6)

30  **$k, k'$  orthogonality**

$$\begin{aligned}\sum_i \psi(i|q, k) \psi^*(i|q, k') &= \sum_i \psi^2(i-qN)E^*(k, i)E(k', i) \\ &= \sum_{i_0} \sum_{\Delta} [\sum_{\Delta q} \psi^2(i_0+\Delta)N] E^*(k, i_0)E(k', i_0) \\ &\cong \sum_{i_0} [1] E^*(k, i_0) E(k', i_0)\end{aligned}$$

$$= N \delta(k-k')$$

using the completeness property of the  
multi-resolution complex Wavelet from reference [2]

$$[\sum_{\Delta} \psi^2(i_0 + \Delta N)] \cong 1 \text{ for all } i_0$$

5  $q, k$  and  $q', k'$  orthogonality proven in reference  
[2] and restated in 4 in equations (6)

3 OWDMA decoding derives estimates  $\hat{x}(k|q)$  of  $x(k|q)$  for  
data block  $q$  from the receiver estimates  $\hat{z}(i)$  of  $z(i)$

$$\begin{aligned} 10 \quad \hat{x}(k|q) &= \text{FWT}^{-1}[\hat{z}(i)] \\ &= \sum_i \hat{z}(i) \psi^*(i|q, k) \end{aligned}$$

4 FWT algorithm for OWDMA decoding in receiver

4.a) FWT pre-sum

$$\begin{aligned} 15 \quad \lambda(i_0|q) &= \sum_{\Delta} \hat{z}(i_0|q+\Delta) \psi(i_0+\Delta N) \\ &= \sum_{\Delta} \hat{z}(i_0+\Delta N|q) \psi(i_0+\Delta N) \end{aligned}$$

4.b) FWT pre-sum FFT

$$\hat{x}(k|q) = \sum_{i_0} \lambda(i_0|q) E(k, i_0)$$

20 5 Computational complexity of fast algorithm

Real multiply rate  $R_M$

$$R_M T = 2 \log_2(N) + 2 L$$

Real add rate  $R_A$

$$R_A T = 3 \log_2(N) + 2 L$$

25

Step 4 is the new fast algorithm for the received OWDMA  
which partitions the baseband symbol detection  $\hat{x}(k|q)$  in step 3  
into a pre-sum calculation sub-step 4.a of  $\lambda(i_0|q)$  followed by a  
sub-step 4.b FFT of this pre-sum.

30

Step 5 evaluates the real multiply complexity metric  $R_M T$  and  
real add computational complexity metrics  $R_A T$  in terms of  
multiplies/adds per digital sample for the fast algorithm in step

3. The first term in these metrics is the complexity of the  $\text{FFT}^{-1}$  for a base 2 and the second term is the complexity of extending the multi-resolution complex Wavelet waveform over  $L$  of the symbol intervals.

5

MS-CDMA parameters and codes are defined in equations (11). Step 1 defines the scenario parameters. Step 2 partitions the user index  $u$  field into the sub-fields  $u_0, u_1$  of size  $N_0, N_1$  for scales 0,1 respectively and which are the indices over the users within each channel and the indices over the channels within the MS-CDMA group and which uniquely represent  $u$  as  $u = u_0 + u_1 N_1$ .

10

Step 3 partitions the code chip index  $n$  field into the sub-fields  $n_0, n_1$  of size  $N_0, N_1$  for scales 0,1 respectively and which are the indices over the chips within each channel and the indices over the channels of the MS-CDMA group and which uniquely represent  $n$  as  $n = n_0 + n_1 N_1$ .

15

Step 4 defines the  $N_c \times N_c$  MS-CDMA code matrix  $C$  whose elements are  $C(u, n)$  where  $u+1$  is the row index and  $n+1$  is the column index and where the  $+1$  has been added to correspond to the row and column numbering starting with  $+1$ . MS-CDMA code vector

20

MS-CDMA parameters and codes (11)

25

1 Scenario parameters

$M$  = number of communications channels  
of band  $B$

$N_0$  = number of CDMA chips per channel

$N_1$  = number of channels in MS-CDMA group

30

$N_c$  = number of chips in MS-CDMA group =  $N_0 N_1$

$1/T_0$  = MS-CDMA chip/symbol rate

=  $1/NT$  for OFDMA, OWDMA in FIG. 1,2

$x(u, q)$  = User  $u$  in group  $q$

=  $x(k|q)$  when  $N_0=1$ ,  $u=k$

$$= x(k) \quad \text{when } N_0=1, u=k, q=0$$

2 User index  $u$  for a MS-CDMA group

$$\begin{aligned} u &= \text{Index of CDMA users} \\ &= 0, 1, \dots, (N_c-1) \\ &= (u_0, u_1) \quad \text{field representation of } u \\ &= u_0 + u_1 N_0 \quad \text{for fields } u_0, u_1 \end{aligned}$$

where the index fields are

$$\begin{aligned} u_0 &= \text{index of users in a channel in} \\ &\quad \text{the MS-CDMA group} \\ &= 0, 1, \dots, (N_0-1) \end{aligned}$$

$$\begin{aligned} u_1 &= \text{index of channels in the MS-CDMA group} \\ &= 0, 1, \dots, N_1-1 \end{aligned}$$

3 Code chip index  $n$  for a MS-CDMA group

$$\begin{aligned} n &= \text{Index of CDMA code chips} \\ &= 0, 1, \dots, (N_c-1) \\ &= (n_0, n_1) \quad \text{field representation of } n \\ &= n_0 + n_1 N_0 \quad \text{for fields } n_0, n_1 \end{aligned}$$

where the index fields are

$$\begin{aligned} n_0 &= \text{index of code chips within a channel} \\ &\quad \text{in the MS-CDMA group} \\ &= 0, 1, \dots, (N_0-1) \end{aligned}$$

$$\begin{aligned} n_1 &= \text{index of channels in the MS-CDMA group} \\ &= 0, 1, \dots, (N_1-1) \end{aligned}$$

4 MS-CDMA code matrix  $C$  for a MS-CDMA group

$$\begin{aligned} C &= N_c \times N_c \text{ MS-CDMA code matrix} \\ &= [C(u, n)] \quad \text{matrix of elements } \{C(u, n)\} \\ &= [c(u)] \quad \text{matrix of } 1 \times N_c \text{ code vectors} \end{aligned}$$

$$\begin{aligned} C(u) &= \text{MS-CDMA code vector } u \text{ which is row } u+1 \text{ in} \\ &\quad C \text{ counting rows, columns starting with 1} \\ &= [C(u, 0), C(u, 1), \dots, C(u, N_c-1)] \end{aligned}$$



## 5 MS-CDMA code

matrix C construction

### 5.a Non-factorable C: algebraic field construction

$$C = [ C(u, n) ]$$

$$= [ C(u_0 + u_1 N_0, n_0 + n_1 N_0) ]$$

### 5.b Factorable C: Tensor product (Kronecker) construction

$$C = C_1 \otimes C_0$$

$$= [ C_1(u_1, n_1) C_0(u_0, n_0) ]$$

where

$$\otimes = \text{Kronecker (tensor) product of } C_1 \text{ and } C_0$$

$$C_1 = N_1 \times N_1 \text{ MS-CDMA scale 1 code matrix}$$

$$= [ C_1(u_1, n_1) ] \text{ matrix of elements } \{ C_1(u_1, n_1) \}$$

$$\text{over the user channels within the group}$$

$$C_0 = N_0 \times N_0 \text{ MS-CDMA scale 0 code matrix}$$

$$= [ C_0(u_0, n_0) ] \text{ matrix of elements } \{ C_0(u_0, n_0) \}$$

$$\text{over the user chips within each channel}$$

$$\text{in the group}$$

$c(u)$  for user  $u$  is the row  $u+1$  of  $C$ . The  $C$  is a complex orthogonal code matrix.

Step 5 illustrates the construction of MS-CDMA code matrix  $C$  for non-factorable and factorable  $C$ . In 5.a for a non-factorable  $C$  the algebraic fields for code indices  $u$  and chip indices  $n$  is constructed as the sum of the scale "0" algebraic fields of the code indices  $u_0$  and chip indices  $n_0$  of users within each channel in the MS-CDMA group, plus the scaled addition of the scale "1" algebraic fields of the code indices  $u_1$  and chip indices  $n_1$  of users over the channels of the MS-CDMA group wherein the scale factor is the size  $N_0$  of the scale "0" algebraic fields of indices. For the special case where  $C$  is factorable in

5.b the  $C$  is constructed as is a Kronecker or Tensor product  $C=C_1 \otimes C_0$  of  $C_1$  and  $C_0$  where " $\otimes$ " is the Kronecker or tensor product and the matrix  $C_1$  is the  $N_1 \times N_1$  MS-CDMA scale "1" code matrix over the user channels within the MS-CDMA group and the matrix  $C_0$  is the  $N_0 \times N_0$  MS-CDMA scale "0" code matrix for the user chips within each channel in the MS-CDMA group.

MS-CDMA representative application to OFDMA in FIG. 1 uses the candidate architecture which spreads the CDMA over the  $M$  contiguous 48 data channels or 52 contiguous data plus pilot channels and the representative application to OFDMA in FIG. 3 uses the candidate architecture which spreads the CDMA over the  $M$  contiguous OFDMA communications channels across  $B$ .

MS-CDMA partitions the  $M$  channels into channel groups of size  $N_1$  and provides a scale over the chips within the channels and another scale over the channels within this group. Code chip length  $N_c$  for each user in a channel group is equal to  $N_c=N_0N_1$  where  $N_0$  is the number of chips in each channel assigned to scale "0" and  $N_1$  is the number of channels assigned to scale "1" within the group. Each user has a chip scale "0" and a channel scale "1". Chip scale "0" spreads the data over the chips within each channel and channel scale "1" then spreads the channel chips uniformly over the  $N_1$  channels with the result that each user occupies each of the channels within the  $N_1$  channel group.

There could be from 1 to  $M/N_1$  channel groups depending on the architecture and applications. The use of multiples groups  $M/N_1 > 1$  tends to be desirable since the storage requirements and computational complexity are reduced as the number of groups are increased and the spreading advantages within each group tend to saturate as the number  $N_1 > 16$  when the channels within each group are spread across the fullband  $M$  channels.

For  $N_0=1$  there is no CDMA within each channel and the MS-CDMA then spreads the signals over each channel within a group for both OFDMA and OWDMA to function as a means to spread each channel over the fullband  $M$  channels and which may be a desirable architecture when the channels are sufficiently narrow to produce a sufficiently long pulse to counter multipath.

MS-CDMA OFDMA transmitter equations are defined in equations (12) for MS-CDMA. Step 1 gives the parameters and definitions. Step 2 defines the encoding equations for chip  $n_0$  for data block  $q$ . Sub-step 2.a uses the fast code transform developed in references [3],[4] to generate the encoded vector. Sub-step 2.b uses the inverse  $FFT^{-1}$  to construct the transmitter complex baseband signal  $z(i_0|n_0+qN_0)$  for chip  $n_0$  for data block  $q$  and these signals are combined in sub-step 2.c to generate the transmitter signal  $z(i)$  for all  $n_0, q$ .

Step 3 evaluates the real multiply complexity metric  $R_{MT}$  and real add computational complexity metrics  $R_{AT}$  in terms of multiplies/adds per digital sample for the fast algorithm in step 3. The first term in these metrics is the complexity of the  $FFT^{-1}$

MS-CDMA OFDMA transmitter equations (12)

1 Parameters and definitions are defined in 1 in equations ( 2),(9),(11) together with

MS-CDMA group index field  $\{g\}$

$g = 0, 1, \dots, (M/N_1-1)$  where  $M/N_1 =$  number of MS-CDMA groups over the  $M$  data channels

Frequency harmonic  $k$

$k = k(n_1|g, q)$

MS-CDMA chip index fields

$n = n_0 + n_1N_0$   
 $= 0, 1, 2, \dots, N_c-1$

$n_0 = 0, 1, 2, \dots, N_0-1$  scale 0 chips in channel

$n_1 = 0, 1, 2, \dots, N_1-1$  scale 1 chips over channels

$N_c = N_0 N_1$  number of chips in a MS-CDMA group

MS-CDMA channel code index  $n_1$

$n_1 = n_1(k|g, q)$  function of  $k$  for a given  $g, q$

User symbol  $x$

$$\begin{aligned} x &= x(u|g, q) \\ &= A(u|g, q) \exp j\phi(u|g, q) \end{aligned}$$

MS-CDMS code orthogonality

$$CC^* = N_c I$$

PN spreading code  $P(n)$  for chip  $n$  in group  $g$

$$P(n|g, q) = \exp(j\phi(n|g, q))$$

Assume the band and pulse filtering can be

neglected on both Tx and Rx which is

equivalent to the assumption  $p(i) \cong 1$  for all  $i$

2 MS-CDMA OFDMA encoding for complex baseband signal  $z(i)$

2.a) FCT MS-CDMA encoding for data block  $q$

$$\begin{aligned} Z(n|g, q) &= \text{FCT}[x(u|g, q)] \\ &= N_c^{-1} \sum_u x(u|g, q) C(u, n) P(n|g, q) \end{aligned}$$

where FCT is the fast code transform

20 2.b)  $\text{FFT}^{-1}$  OFDMA encoding for chip  $n_0$  of data block  $q$

$$\begin{aligned} z(i_0|n_0+qN_0) &= \text{FFT}^{-1}[Z(n|g, q)] \\ &= N^{-1} \sum_k Z(n|g, q) E^*(k, i_0) \end{aligned}$$

where  $n = n_0 + n_1 N_0$

$$n_1 = n_1(k|g, q)$$

$$Z(n|g, q) = Z(n_0 + n_1(k|g, q)N_0)$$

25 2.c) Complex baseband transmitter signal  $z(i)$

$$z(i) = \sum_{(o)} z(i_0|n_0+qN_0) \quad \text{for } (o) = (n_0+qN_0)N$$

3 Computational complexity

30 Real multiply rate  $R_M$

$$R_M T = 2 \log_2(N)$$

Real add rate  $R_A$

$$R_A T = 3 \log_2(N) + 2 \log_2(N_c) + 2$$

for a base 2 and the second term is the complexity of the FCT assuming the FCT does not require any multiplies.

MS-CDMA OWDMA transmitter equations are defined in  
 5 equations (13) for MS-CDMA. Step 1 lists the parameters and definitions. Step 2 defines the encoding equations for chip  $n_0$  for data block  $q$ . The FCT on the symbols in sub-step 2.a yields the encoded data block  $Z(n|g,q)$  and the FWT on this output in sub-step 2.b yields the transmitter complex baseband signal  
 10  $z(i_0|n_0+qN_0)$  for chip  $n_0$  for data block  $q$  and these signals are combined in sub-step 2.c to generate the transmitter signal  $z(i)$  for all  $n_0, q$ .

MS-CDMA OWDMA transmitter equations (13)

15 1 Parameters and definitions are defined in 1 in equations (9), (12) together with the range of  $q$   
 $q = 0$  for the data block being addressed  
 $= +/-1$  contiguous data blocks  
 Assume the  $h(f)$  is flat over the passband for  
 20 both Tx and Rx and can be neglected

2 MS-CDMA OWDMA encoding for chip  $n_0$  for data block  $q$

2.a) FCT MS-CDMA encoding for data block  $q$

$$Z(n|g,q) = \text{FCT}[x(u|g,q)]$$

25  $= N_c^{-1} \sum_u x(u|g,q) C(u,n) P(n|g)$

2.b) FWT OWDMA encoding for chip  $n_0$  of data block  $q$

$$z(i_0|n_0+qN_0) = \sum_{\delta q} \sum_{\Delta} \text{FWT}[Z(n+\Delta|g,q+\delta q)]$$

$$= N^{-1} \sum_k \sum_{\delta q} \sum_{\Delta} Z(n+\Delta|g,q+\delta q) \psi(i|n_0+\Delta, k)$$

$$= N^{-1} \sum_k \sum_{\delta q} \sum_{\Delta} Z(n+\Delta|g,q+\delta q) \psi(i_0+\Delta N) E^*(k, i_0)$$

30 where the overlap conditions on  $\delta q$  are

$$\delta q = \Delta \text{ for } N_0=1$$

$$= \text{sign}(\Delta+n_0) \lfloor |\Delta+n_0|/N_0 \rfloor \text{ otherwise}$$

$$\text{sign}(o) = \text{numerical sign of } (o)$$

$\lfloor (o) \rfloor$  = rounded down integer value of  $(o)$

where the overlap index  $\Delta = 0, +/-1, .$

$. . . , +/- (L/2-1)$

2.c) Complex baseband transmitter signal  $z(i)$

$$z(i) = \sum_{(o)} z(i_o | n_o + qN_o) \quad \text{for } (o) = (n_o + qN_o)N$$

3 FWT algorithm for OWDMA encoding in the transmitter

3.a) FWT pre-calculation  $FFT^{-1}$

$$\lambda(i_o | n_o + \Delta, q)$$

$$= N^{-1} \sum_k Z(n + \Delta | g, q) E^*(k, i_o)$$

$$= FFT^{-1}[N^{-1} \sum_k Z(n + \Delta | g, q) E^*(k, i_o)]$$

**3.b) FWT post-sum**

$$z(i_o | n_o) = \sum_q \sum_{\Delta} \psi(i - (n_o + \Delta)N) \lambda(i_o | n_o + \Delta, q)$$

4 Computational complexity for fast algorithms in 2.a,3

Real multiply rate  $R_M$

$$R_{MT} = 2 \log_2(N) + 2L$$

Real add rate  $R_A$

$$R_{AT} = 3 \log_2(N) + 2L + 2 \log_2(N_c) + 2$$

Step 3 is the new fast FWT algorithm in this invention disclosure for the transmitted OWDMA which consists of the pre-calculation  $FFT^{-1}$  in sub-step 3.a followed a post-sum in sub-step 3.b of the product from sub-step 3.a with the corresponding

Wavelet overlaps over the  $q$  data block interval.

Step 4 evaluates the real multiply complexity metric  $R_{MT}$  and real add computational complexity metrics  $R_{AT}$  in terms of multiplies/adds per digital sample for the fast algorithms FCT and FWT in steps 2,3. The first term in these metrics is the complexity of the  $FFT^{-1}$  for a base 2, the second term  $2L$  is the complexity of extending the multi-resolution complex Wavelet waveform over  $L$  of the MT symbol intervals, and the remaining

terms are the complexity of the FCT assuming the FCT does not require any multiplies.

MS-CDMA OFDMA/OWDMA encoding for the transmitter in FIG. 6 is a representative implementation of the MS-CDMA OFDMA encoding algorithms in equations (12) and the MS-CDMA OWDMA encoding algorithms in equations (13). Signal processing starts with the input stream of data encoded symbols  $x(u|g,q)$  40 from the transmitter symbol encoder 52 in FIG. 7A and defined in 1 in equations (12) for both OFDMA and OWDMA.

FIG. 6 MS-CDMA encoding 41 implements the fast code transform FCT encoding defined in sub-step 2.a in equations (12), (13) for the MS-CDMA encoding and PN cover or spreading encoding to generate the  $N_c$  outputs  $x(u|g,q)C(u,n)P(n|g)$  for each MS-CDMA group  $g$  and data block  $q$ . For each group these outputs are summed 42 over  $u$  to generate the encoded vector  $Z(n|g,q)$  in 43.

OFDMA processing 44 performs an  $FFT^{-1}$  on the received set of vectors  $Z(n|g,q)$  and a summation to implement sub-steps 2.b, 2.c in equations (12) and the output is band filtered 46 to generate the MS-CDMA OWDMA encoded complex baseband signal  $z(i)$  in 47.

OWDMA processing 45 performs an FWT on the received set of vectors  $Z(n|g,q)$  and a summation to implement sub-steps 2.b, 2.c in equations (13) and the output is band filtered 46 to generate the MS-CDMA OWDMA encoded complex baseband signal  $z(i)$  in 47.

Outputs  $z(i_0|n_0)$  47 from the MS-CDMA OFDMA and MS-CDMA OWDMA are digital-to-analog DAC converted 48 and handed off to the analog front end 49 as the complex baseband analog signal  $z(t)$  in 49.

MS-CDMA OFDMA/OWDMA transmitter description in FIG. 7 presents a block diagram in FIG. 7A and a representative MS-CDMA mapping in FIG. 7B. FIG. 7A is a representative transmitter implementation of the MS-CDMA OFDMA and MS-CDMA OWDMA encoding in FIG. 6. The transmitter block diagram in FIG. 7A includes the FIG. 6 MS-CDMA OFDMA/OWDMA encoding in an abbreviated format 54 in FIG. 7A. FIG. 7A signal processing starts with the stream of user input data words. Frame processor 51 accepts these data words and performs the encoding and frame formatting wherein CRC is a cyclic redundant code for error detection, and passes the outputs to the symbol encoder 52 which encodes the frame symbols into amplitude (Ampl.) and phase coded symbols  $x(u|g,q)$  53 which are the input to the MS-CDMA encoding 55 and which is 41,42 in FIG. 6.

MS-CDMA FCT encoding outputs  $Z(n|g,q)$  56 are handed over to the OFDMA and OWDMA processing 57 which performs an inverse  $FFT^{-1}$  followed by a band filtering for OFDMA which is 44,46 in FIG. 4 and performs an inverse  $FWT^{-1}$  followed by a band filtering which is 45,46 in FIG.6. This complex baseband signal  $z(i)$  in 47 in FIG.6 is digital-to-analog DAC converted 59 and the output complex baseband analog signal  $z(t)$  60 is handed off to the analog front end 61.

The  $z(t)$  is single sideband upconverted, amplified, and transmitted (Tx) by the analog front end 61 as the real waveform  $v(t)$  62 at the carrier frequency  $f_0$  whose amplitude is the real part of the complex envelope of the baseband waveform  $z(t)$  and the phase angle  $\phi$  accounts for the phase change from the baseband signal to the transmitted signal. Output waveform 62 from the analog front end is the Tx waveform from the Tx antenna.

FIG. 7B illustrates a representative MS-CDMA uniform mapping of each data symbol over frequency, time, antennas, and



beams of a cellular communications transmitter. Multiple antennas and beams are used when a multiple-input-multiple-output MIMO communications link is being implemented. The algebraic field construction of the algebraic index fields for the codes and chips for a 2-scale MS-CDMA construction of a non-factorable code matrix  $C$  in equations (11) represented by 151 and 152 in FIG. 7B for the algebraic chip indices  $n_0, n_1$  is continued in FIG. 7B to include the algebraic chip index fields for chips  $n_2$  in 153 over the frequency bands, chips  $n_3$  in 154 over the data blocks, chips  $n_4$  in 155 over the transmit antenna beams, and chips  $n_5$  in 156 over the transmit antennas. The corresponding algebraic code indices are respectively  $u_0, u_1, u_2, u_3, u_4, u_5$  and the MS-CDMA code length is  $N_c = N_0 N_1 N_2 N_3 N_4 N_5$  chips.

It should be obvious to anyone skilled in the communications art that this example implementation in FIG. 6,7 clearly defines the fundamental MS-CDMA OFDMA and MS-CDMA OWDMA signal processing relevant to this invention disclosure and it is obvious that this example is representative of the other possible signal processing approaches.

MS-CDMA OFDMA receiver equations are defined in equations (14). Step 1 lists the parameters and definitions and the assumption that the band and pulse filtering can be neglected. Step 2 defines decoding of the received chip signal to derive the estimate  $\hat{x}(u|g,q)$  of the transmitted symbol  $x(u|g,q)$ . Sub-step 2.a derives the estimate for the encoded symbols using the FFT and sub-step 2.b uses the  $FCT^{-1}$  on this estimate to derive the transmitted symbol estimate.

MS-CDMA OFDMA receiver equations (14)

1 Parameters and definitions are defined in 1 in equations (2),(9),(11) together with

$\hat{z}(i)$  = receiver band filter estimate of

the transmitted complex baseband signal  $z(i)$   
 Assume the band and pulse filtering can be  
 neglected on both Tx and Rx which is  
 equivalent to the assumption  $p(i) \cong 1$  for all  $i$

5

2 MS-CDMA OFDMA decoding to derive data symbol  $\hat{x}(u|g,q)$

2.a) FFT of each received chip vector  $\hat{z}(i_0|n_0+qN_0)$

$\hat{Z}(n|g,q)$  = estimate of transmitted  $Z(n|g,q)$

$$= \text{FFT}[ \hat{z}(i_0|n_0+qN_0) ]$$

10

$$= \sum_{i_0} \hat{z}(i_0|n_0+qN_0) E(k, i_0)$$

where  $\hat{z}(i_0|n_0+qN_0) = \hat{z}(i=i_0+(n_0+qN_0)N)$

2.b)  $\text{FCT}^{-1}$  of each group  $g$  encoded vector  $Z(n|g,q)$

$$\hat{x}(u|g,q) = \text{FCT}^{-1}[ \hat{Z}(n|g,q=0) ]$$

$$= \sum_n \hat{Z}(n|g,q) P^*(n|g) C^*(u,n)$$

15

3 Computational complexity is the same as calculated  
 in 3 in equations (12) for the transmitter

MS-CDMA OFDMA receiver equations are defined in equations  
 (15). Step 1 lists the parameters and definitions and the  
 assumption that the band and pulse filtering can be neglected.

20

MS-CDMA OFDMA receiver equations (15)

1 Parameters and definitions are defined in 1 in

25

equations (2), (9), (11), (14) together with

$\hat{z}(i)$  = receiver band filter estimate of

the transmitted complex baseband signal  $z(i)$

Assume the  $h(f)$  is flat over the passband for

both Tx and Rx and can be neglected

30

2 MS-CDMA OFDMA decoding to derive data symbol  $\hat{x}(u|g,q)$

2.a)  $\text{FWT}^{-1}$  of each received signal  $\hat{z}(i)$

$$\hat{Z}(n|g,q) = \text{estimate of transmitted } Z(n|g,q)$$

$$\begin{aligned}
&= \text{FWT}^{-1}[\hat{z}(i)] \\
&= \sum_i \hat{z}(i) \psi^*(i|q, k) \\
&= \sum_{i_0} \sum_{\Delta} \hat{z}(i_0|n_0+(q+\Delta)N) \psi(i_0+\Delta N) E(k, i_0)
\end{aligned}$$

where  $\hat{z}(i_0|n_0+qN_0) = \hat{z}(i=i_0+(n_0+qN_0)N)$

5  $k \rightarrow g$  which means  $k$  specifies  $g$   
 $n_0, k \rightarrow n=n_0+n_1(k|g, q)$

2.b)  $\text{FCT}^{-1}$  of each group  $g$  encoded vector  $Z(n|g, q)$

$$\begin{aligned}
\hat{x}(u|g, q) &= \text{FCT}^{-1}[\hat{Z}(n|g, q)] \\
10 \quad &= \sum_n \hat{Z}(n|g, q) P^*(n|g) C^*(u, n)
\end{aligned}$$

3  $\text{FWT}^{-1}$  algorithm

2.a)  $\text{FWT}^{-1}$  pre-sum

$$\lambda(i_0|g, q) = \sum_{\Delta} \hat{z}(i_0|n_0+(q+\Delta)N) \psi(i_0-(n_0+\Delta)N)$$

15 where  $n_0+\Delta \rightarrow g$  using the boundary conditions  
in 2.b in equations (13)

2.b)  $\text{FWT}^{-1}$  pre-sum FFT

$$\begin{aligned}
\hat{Z}(n|g, q) &= \sum_{i_0} \text{FFT}[\lambda(i_0|g, q)] \\
&= \sum_{i_0} \lambda(i_0|g, q) E(k, i_0)
\end{aligned}$$

20

3 Computational complexity is the same as calculated  
in 3 in equations (13) for the transmitter

Step 2 defines decoding of the received chip signal to derive  
25 the estimate  $\hat{x}(u|g, q)$  of the transmitted symbol  $x(u|g, q)$ . Sub-  
step 2.a derives the estimate for the encoded symbols using the  
 $\text{FWT}^{-1}$  and sub-step 2.b uses the  $\text{FCT}^{-1}$  on this estimate to derive  
the transmitted symbol estimate. Step 3 is the new fast FWT  
algorithm in this invention disclosure for the received MS-CDMA  
30 OWDMA which consists of the pre-sum in sub-step 3.a followed an  
FFT on this pre-sum.

MS-CDMA OFDMA/OWDMA decoding for the receiver in FIG. 8 is a representative implementation of the MS-CDMA OFDMA decoding algorithms in equations (14) and the MS-CDMA OWDMA decoding algorithms in equations (15). Signal processing starts with the  
5 input intermediate frequency IF signal after being single sideband downconverted and synchronized in frequency 63 from the receiver front end 72 in FIG. 9. This input signal is band filtered 64 and handed off to the analog-to-digital converter ADC or A/D 65 whose digital output is the received estimate  $\hat{z}(i)$  66  
10 of the transmitted complex baseband signal  $z(i)$  60 in FIG. 7. For OFDMA 67 this received signal is processed by an FFT to derive estimates  $\hat{Z}(n|g,q)$  68 of the MS-CDMA encoded signal  $Z(n|g,q)$  56 in FIG. 7. For OWDMA 67 this received signal is processed by an inverse multi-resolution complex Wavelet  
15 transform  $\text{FWT}^{-1}$  to derive estimates  $\hat{Z}(n|g,q)$  68 of the MS-CDMA encoded signal  $Z(n|g,q)$  56 in FIG. 7 implementing the fast algorithm defined in 2 in equations (15). Recovered estimates  $\hat{Z}(n|g,q)$  68 are processed by the inverse fast code transform  $\text{FCT}^{-1}$  69 to derive estimates  $\hat{x}(u|g,q)$  70 of the transmitted data  
20 symbols  $x(u|g,q)$  53 in FIG. 7 for hand over to the symbol decoder.

MS-CDMA OFDMA/OWDMA receiver block diagram in FIG. 9 is a representative receiver implementation of the MS-CDMA OFDMA and  
25 MS-CDMA OWDMA decoding in FIG. 8. The receiver block diagram in FIG. 9 includes the FIG. 8 MS-CDMA OFDMA/OWDMA decoding in an abbreviated format 74 in FIG. 9. FIG. 9 signal processing starts with the received Rx waveform 71 from the transmitter 62 in FIG. 7. Received (Rx) signal  $\hat{v}(t)$  71 is an estimate of the transmitted  
30 signal  $v(t)$  62 in FIG. 7 received with errors in time  $\Delta t$ , frequency  $\Delta f$ , phase  $\Delta \theta$ , and with an estimate  $\hat{z}(i)$  of the transmitted complex baseband signal  $z(t)$  60 in FIG. 7. This received signal  $\hat{v}(t)$  is amplified and downconverted by the analog

front end 72 and single side band SSB downconverted and  
synchronized 73 and handed over to the digital-to-analog  
conversion DAC processing 75 for band filtering by  $h(f)$  and  
digitization to generate the baseband signal  $\hat{z}(i)$  76 which is the  
5 received estimate of the transmitted signal  $z(i)$  60 in FIG. 7.  
Timing synchronization could be implemented in the DAC.

Outputs  $\hat{z}(i)$  are processed by the MS-CDMA OFDMA/OWDMA  
decoding to derive estimates  $\hat{x}(u|g,q)$  79 of the transmitted  
10 symbols  $x(u|g,q)$  53 in FIG. 7 and part of the information is  
handed off to the synchronization (sync) processor. For the  
inverse MS-CDMA<sup>-1</sup> OFDMA<sup>-1</sup> the processing 79 consists of the FFT to  
recover estimates  $\hat{Z}(n|g,q)$  of  $Z(n|g,q)$  followed by an inverse  
FCT<sup>-1</sup> to recover  $\hat{x}(u|g,q)$ . For the inverse MS-CDMA<sup>-1</sup> OWDMA<sup>-1</sup> the  
15 processing 79 consists of the inverse FWT<sup>-1</sup> to recover estimates  
 $\hat{Z}(n|g,q)$  of  $Z(n|g,q)$  followed by an inverse FCT<sup>-1</sup> to recover  
 $\hat{x}(u|g,q)$  and the fast transform in 2 in equations (15). Outputs  
 $\hat{x}(u|g,q)$  are processed by the symbol decoder 80 and the frame  
processor 81 for handoff as the received Rx data 82.

20

It should be obvious to anyone skilled in the  
communications art that this example implementation in FIG. 8,9  
clearly defines the fundamental MS-CDMA OFDMA and MS-CDMA OWDMA  
signal processing relevant to this invention disclosure and it  
25 is obvious that this example is representative of the other  
possible signal processing approaches.

Variable power control across the frequency band can be  
implemented by assigning each group  $g$  of transmit Tx signals  
30 their own power level  $P(g)$ . Each MS-CDMA group  $g$  occupies a  
subset of the channels over the frequency band  $B$  consisting of  
 $N_1$  channels which means that the users within group  $g$  are  
transmitted with the same Tx power. On receive each group  $g$  of  
channels is processed separately so there is no cross-talk

between the users in the different groups. The OWDMA was designed to support large dynamic range imbalances between channels which could be present with power level control. MS-CDMA OFDMA/OWDMA variable power control over the frequency subbands corresponding to the MS-CDMA groups supports the potential for diversity improvements by allocation of the available power to emphasize the 'best' set of available subbands, which subbands are not required to be contiguous as well as supporting the simultaneous support of multiple users with differing power requirements due to range, multi-path, and path attenuation effects.

A second configuration for variable power control is described in reference [2] and increases the flexibility of power control to all of the individual channels.

Preferred embodiments in the previous description is provided to enable any person skilled in the art to make or use the present invention. The various modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without the use of the inventive faculty. Thus, the present invention is not intended to be limited to the embodiments shown herein but is not to be accorded the wider scope consistent with the principles and novel features disclosed herein.

30